Annals of the Academy of Romanian Scientists
Series on Science and Technology of Information
ISSN 2066 - 8562 Volume **4**, Number **1**/**2011**

BASEBAND BLOCKS NOISE PARTITIONING IN MULTI−STANDARD WIRELESS RECEIVERS EMBEDDING ANALOG SIGNAL CONDITIONING

Silvian SPIRIDON[1], Florentina SPIRIDON[1], Claudius DAN[2], Mircea BODEA[2]

Rezumat. *Lucrarea prezintă strategia partiţionării zgomotului între blocurile componente ale parţii de joasă frecvenţă ale unui radio receptor multi-standard cu conversie directă de frecvenţă bazat pe condiţionarea analogică a semnalului. În urma unei analize de prim ording la nivel de sistem, lucrarea construieşte un model de zgomot pentru blocurile componente ale parţii de joasă frecvenţă. Modelul este centrat pe sub-circuitul cu este contruită partea de joasă frecvenţă a receptorului multi-standard: amplificatorul diferenţial cu reacţie negativă. Astfel, contribuţiile individuale ale blocurilor componente ale părţii de joasă frecvenţă ale receptorului sunt calculate. Scopul principal al lucrării este tratarea corespunzătoare a compromisului între consumul de putere şi arie în vederea partiţionării zgomotului între blocurile componente ale parţii de joasă frecvenţă ale receptorului.*

Abstract. *This paper presents the noise partitioning strategy for the Low Frequency (LF) part of Direct Conversion CMOS multi-standard wireless receiver embedding analog baseband signal conditioning. Based on a first order system level analysis, the paper builds a noise model for the receiver LF part blocks. The in-depth circuit level noise analysis centers the model on the LF chain building brick: the fully differential feed-back amplifier embedding a linear feedback network. The baseband noise partitioning is shaped by the trade-off between the LF part active circuits' power consumption and the LF part passive components area. In order to efficiently address this trade-off the paper introduces a new concept: the baseband noise excess factor (k_{LF}). The factor accounts the feed-back amplifier excess opamp noise contribution with respect to its feed-back resistors noise contribution. Thus, by sizing the factor the designer is enabled to proficiently trade-off the between the circuits power consumption and area.*

Keywords: Software Defined Radio, Direct Conversion Receiver, Noise partitioning

1. Introduction

The homodyne quadrature down-converter architecture provides the optimum solution for the implementation of Re-Configurable Multi-Standard Radio Receivers, [1]. In spite recently newer digital assisted techniques have been introduced to allow the reduction of the analog circuitry [2, 3], the most common multi-standard receiver architecture embeds analog signal conditioning.

[1]Ph.D. Student, Electronics, Telecommunications and Information Technology Department, University “Politehnica” of Bucharest, Romania.
[2]Professor, Electronics, Telecommunications and Information Technology Department, University “Politehnica” of Bucharest, Romania.

In [4] the multi-standard receiver architecture embedding analog signal conditioning has been presented and the receiver building blocks have been introduced. Basically, the receiver chain is split into a high frequency part (HF), comprised by the Low Noise Amplifier (LNA) and the g_m stage of the quadrature down-converter mixer (MIX) and a remaining baseband, low-frequency (LF) part, following the mixer's switching stage. After mixing, the signal is conditioned by a channel selection Low Pass Filter (LPF) and a Variable Gain Amplifier (VGA), before its conversion to digital by an analog-to-digital converter (ADC).

Section 2 analysis the receiver LF part building brick: the fully differential low power amplifier embedding a linear feed-back network. In Section 3 the receiver noise partitioning between its HF and LF parts is reveled, while Section 4 introduces the LF part noise excess factor k_{LF}.

Section 5 reveals the trade-off between the LF part blocks power consumption and area; in Section 6 the baseband noise partitioning is completed by giving the same importance to both area and power consumption of the LF circuits. Finally, Section 4 closes the paper by presenting the conclusions.

2. Low Frequency Chain Building Blocks

The main target in designing the CMOS multi-standard radio receiver LF blocks is the development of a modular architecture that ensures design robustness with respect to both the multi-standard environment (i. e. the variable baseband bandwidths) and the easiness of design porting to a smaller feature CMOS process.

The optimal direct conversion receiver design makes the noise of the baseband chain less critical (due to the RF front-end gain), while linearity performance is more stringent (due to lack of consistent filtering on the RF path). Thus, the feedback use represents the only way the designer can control and, subsequently, meet the specifications, alleviating the specifics of the technology implementation, [5].

This aspect becomes more and more critical as the SoC implementation moves towards deep sub-micron CMOS processes, where system level design should not be limited by particular technology characteristics, like leakage, for instance. Hence, all baseband blocks will contain opamps that sustain a feed-back network.

The basic schematic of the baseband chain building blocks is presented in Fig. 1.a, redrawn from [4]. The base amplifier is a fully differential opamp, while the feed-back network is made out of linear elements, like resistors or/and capacitors.

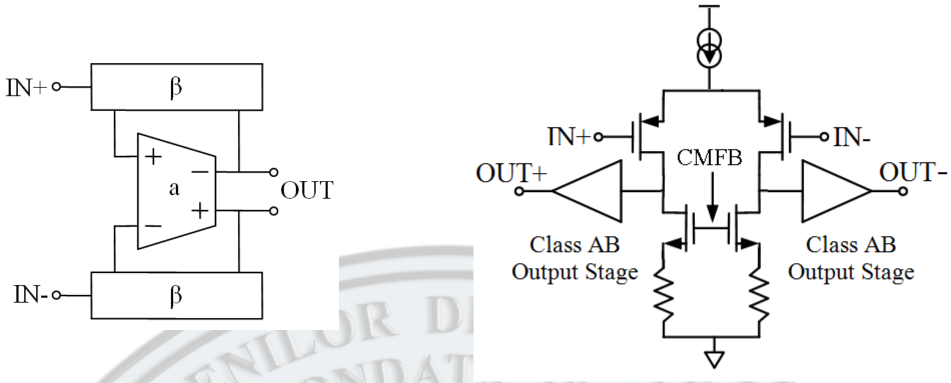

Fig. 1. – a. LF Chain Building Brick and b. Generic Opamp Implementation.

The optimum design of the base amplifier, a fully differential two stage opamp, is analyzed in reference [6]. The opamp generic block diagram is depicted in Fig. 1.b.

The two stage opamp implementation was driven by noise-linearity-power consumption trade-offs. First of all, the stringent noise requirements of wireless standards lead to low values for the feedback resistors, as is detailed in Section 4. Hence, the opamp output stage will act as a buffer, reducing the loading effect on its intrinsic parameters (like GBW) and preserving the intrinsic performance of the first stage.

The opamp 1/*f* noise optimization lead to the choice of the p-channel input stage transistors, while its power consumption optimization sets the class AB topology for the output stage. Also, since the amplifier is fully differential, a common mode feed-back loop (CMFB) is required to set the amplifiers output common mode voltage.

3. Receiver Noise Partitioning Strategy

Today's commercially available receivers have a very small receiver *NF*, NF_{RX}, of about 3 dB, [7]. The 3 dB overall receiver *NF* budget must be partitioned between the multi-standard receiver HF and *LF* parts.

According to Friis equation, [6], the receiver global spot *NF*, NF_{RX}, can be calculated from the individual contributions of the receiver HF and *LF* parts:

$$NF_{\mathrm{RX}} = 10\log\left(F_{\mathrm{HF}} + \frac{F_{\mathrm{LF}} - 1}{A_{V\,\mathrm{HF}}^{2}}\right) dB, \tag{1}$$

where F_{HF}, respectively F_{LF}, represent the spot noise factors of the receiver HF part, respectively *LF* part, and $A_{V\,\mathrm{HF}}$ is the receiver's HF front-end voltage gain.

From eq. (1) becomes clear the RF front-end gain reduces the LF blocks noise contribution. However, the maximum $A_{V\,\mathrm{HF}}$ is limited by linearity constraints. Since the interferers/blockers level present at the receiver input may be considerably larger than the useful signal (e. g. ≥ +70 dBc, [4]), a too large $A_{V\,\mathrm{HF}}$ value can clip the receiver RF front-end output. It also leads to poor linearity for the LNA or the mixer circuits. Thus, typically the $A_{V\,\mathrm{HF}}$ maximum value is limited to 200 or 46 dB. In this paper, for further calculations we shall consider $A_{V\,\mathrm{HF}} = 40$ dB.

Since the LF part noise contribution is strongly reduced by $A_{V\,\mathrm{HF}}$, the HF part is allowed to contribute the most to the overall receiver noise for its power consumption reduction. So, we assume only 1 dB from NF_{RX} is allocated for the LF chain. From it, 0.5 dB must be reserved for the ADC noise.

Thus, it results the input referred noise of the receiver LF part analog baseband blocks corresponds to only 0.5 dB from the 3 dB total NF_{RX} budget. Hence, we can calculate F_{LF} as:

$$F_{\mathrm{LF}} = 1 + A_{V\,\mathrm{HF}}^2 \left(10^{NF_{\mathrm{RX}}/10} - F_{\mathrm{HF}}\right) \cong 4000 \tag{2}$$

4. LF Part Noise Excess Factor

Given the baseband building block structure (see Fig. 1.a), the equivalent input referred noise spectral density at the LF chain input, $\overline{v_{n\,\mathrm{LF}}^2}/\Delta f$, can be split between the noise contributions from the LF operational amplifiers, $\overline{v_{n\,\mathrm{LF\,a}}^2}/\Delta f$, and from the LF part feedback network, $\overline{v_{n\,\mathrm{LF}\,\beta}^2}/\Delta f$:

$$\overline{v_{n\,\mathrm{LF}}^2}/\Delta f = \overline{v_{n\,\mathrm{LF\,a}}^2}/\Delta f + \overline{v_{n\,\mathrm{LF}\,\beta}^2}/\Delta f\,, \tag{3}$$

Thus the LF part of the spot noise factor, F_{LF}, results as:

$$F_{\mathrm{LF}} = 1 + \left(\overline{v_{n\,\mathrm{LF}}^2}/\Delta f\right) / (k_B T R_S)\,, \tag{4}$$

where k_{B} is the Boltzmann constant, T the absolute temperature and R_S the equivalent antenna noise resistance.

By introducing eq. (3) into eq. (4) we can clearly distinguish the two contributions to F_{LF}: one is originating from the feedback resistors and the other one is the overhead generated by the LF part operational amplifiers.

In order to properly asses the overhead to the overall noise budget of the LF part operational amplifiers we introduce the LF part excess noise factor k_{LF}, defined by:

$$k_{\mathrm{LF}} = \left(\overline{v_{n\,\mathrm{LF\,a}}^2}\Big/\Delta f\right)\Big/\left(\overline{v_{n\,\mathrm{LF}\,\beta}^2}\Big/\Delta f\right) \tag{5}$$

Thus, using k_{LF} we can re-write eq. (4) to:

$$F_{LF} = 1 + \left(1 + k_{\mathrm{LF}}\right)\left(\overline{v_{n\,\mathrm{LF}\beta}^2}\Big/\Delta f\right)\Big/\left(k_B T R_S\right), \tag{6}$$

The LF part feedback resistors noise spectral density is equivalent to:

$$\overline{v_{n\,\mathrm{LF}\,\beta}^2}\Big/\Delta f = 4k_B T R_{n\,\mathrm{LF}\,\beta}, \tag{7}$$

where $R_{n\,\mathrm{LF}\,\beta}$ is the LF part equivalent noise resistance of the differential amplifiers feed-back networks.

Similarly the noise spectral density originating from the LF feed-back amplifiers is given by:

$$\overline{v_{n\,\mathrm{LF\,a}}^2}\Big/\Delta f = 4k_B T R_{n\,\mathrm{LF\,a}}, \tag{8}$$

where $R_{n\,\mathrm{LF\,a}}$ represents the equivalent noise resistance of the LF baseband amplifiers.

The LF part opamp noise performance and $R_{n\,\mathrm{LF\,a}}$ versus opamp power consumption dependency were evaluated in reference [8].

Given the noise contributions from eqs. (7) and (8), the total input referred LF noise spectral density of eq. (3) becomes:

$$\overline{v_{n\,\mathrm{LF}}^2}\Big/\Delta f = 4k_B T\left(R_{n\,\mathrm{LF\,a}} + R_{n\,\mathrm{LF}\,\beta}\right), \tag{9}$$

Thus the LF part spot noise factor, F_{LF}, results as:

$$F_{LF} = 1 + \left(1 + k_{\mathrm{LF}}\right) 4R_{n\,\mathrm{LF}\,\beta}/R_S \tag{10}$$

5. The Trade-off between the LF Part Power Consumption and Area

From eq. (10), the minimum F_{LF}, F_{LFmin}, is achieved when the base amplifiers noise is negligible compared to the feedback resistance noise ($k_{\mathrm{LF}} = 0$). By rearranging eq. (10), the $R_{n\,\mathrm{LF}\,\beta}/R_S$ ratio as a function of k_{LF} results as:

$$\frac{R_{n\,\mathrm{LF}\beta}}{R_S} = \frac{F_{\mathrm{LF}} - 1}{4\left(1 + k_{\mathrm{LF}}\right)} \tag{11}$$

Fig. 2.a plots the $R_{n\,\mathrm{LF}\,\beta}/R_S$ ratio as a function of k_{LF}. As expected, the larger the noise spectral density from the base opamps, a smaller value for feed-back resistors is required to keep the same noise spectral density contribution for the baseband blocks.

But, in the same time, the integrated capacitance must be increased accordingly to the requirement of maintaining the low pass filter (LPF) bandwidth and, thus, the baseband integrated output noise. This is the key trade-off between the LF circuit's power consumption and area.

Fig. 2.b plots on the same graph $R_{n\,\mathrm{LF}\,\beta}$, as a measure of the area consumption, and $R_{n\,\mathrm{LF}\,a}$, as a measure of the power consumption, versus k_{LF} for $R_S = 100\ \Omega$.

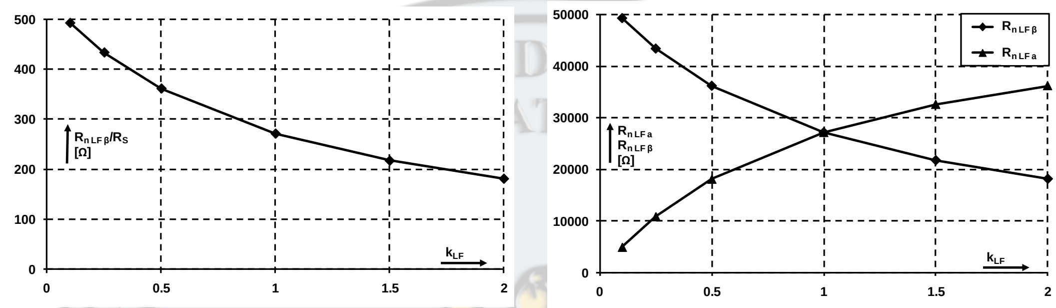

Fig. 2. – a. R_n / R_S and b. R_n, $G_{mn\,\mathrm{LF}}$ vs. k_{LPF}.

Due to the low bandwidth of the envisaged standards (i.e. 100 kHz for GSM), the receiver area is mainly determined by the amount of integrated LPF capacitance. Hence, the trade-off represented in Fig 2.b is the key issue of the receiver baseband noise partitioning.

6. Baseband Noise Partitioning

The LF chain analog building blocks are the mixer transimpedance amplifier, the LPF and the VGA. Since all of these blocks are based either on one or on a cascaded series of fully differential feed-back amplifiers embedding linear feed-back networks (e. g. mixer – [4], LPF – [9], and VGA – [10]), the noise contribution of the individual blocks can be split between the base operational amplifiers and the feed-back network resistors. The noise breakdown of the LF part blocks is presented in Table 1.

Given the notations from Table 1, the total input referred LF noise spectral density is:

$$\begin{aligned} \overline{v_{n\,\mathrm{LF}}^2}\big/\Delta f &= \overline{v_{n\,\mathrm{MIX}}^2}\big/\Delta f + \overline{v_{n\,\mathrm{LPF}}^2}\big/\Delta f + \overline{v_{n\,\mathrm{VGA}}^2}\big/\Delta f = \\ &= 4k_B T \times (1 + k_{\mathrm{LF}}) R_{n\,\mathrm{LF}} = \\ &= 4k_B T \times (1 + k_{\mathrm{LF}}) (R_{n\,\mathrm{MIX}} + R_{n\,\mathrm{LPF}} + R_{n\,\mathrm{VGA}}) \end{aligned} \tag{12}$$

where k_{MIX}, k_{LPF} and k_{VGA} represent the mixer, LPF and VGA noise excess factors.

In line with the modular design of the receiver LF part, its noise partitioning assumes $k_{\mathrm{LF}} = k_{\mathrm{MIX}} = k_{\mathrm{LPF}} = k_{\mathrm{VGA}}$.

In order to optimize the area consumption, not all of the three individual blocks noise contributions will be the same. Since the receiver area is mainly dominated by the amount of integrated LPF capacitance, the LPF will be allowed to contribute as much as the mixer transimpedance amplifier and the VGA all together. This translates to the following condition:

$$R_{n\,\mathrm{LPF}\beta} = 2R_{n\,\mathrm{MIX}\,\beta} = 2R_{n\,\mathrm{VGA}\,\beta} = R_{n\,\mathrm{LF}\,\beta}/2 \quad (13)$$

Table 1. Receiver LF Part Noise Breakdown

LF Block	Contributors	Formula	Notes
Mixer Baseband Amplifier	Base amplifier	$\overline{v_{n\,\mathrm{MIX\,a}}^2} = 4k_B T R_{n\,\mathrm{MIX\,a}} \Delta f$	$R_{n\,\mathrm{MIX\,a}}$ is the mixer's opamp equivalent noise resistance
	Feedback network	$\overline{v_{n\,\mathrm{MIX}\,\beta}^2} = 4k_B T R_{n\,\mathrm{MIX}\,\beta} \Delta f$	$R_{n\,\mathrm{MIX}\,\beta}$ is the mixer's feedback network equivalent noise resistance
	TOTAL	$\dfrac{\overline{v_{n\,\mathrm{MIX}}^2}}{\Delta f} = \dfrac{\overline{v_{n\,\mathrm{MIX\,a}}^2}}{\Delta f} + \dfrac{\overline{v_{n\,\mathrm{MIX}\,\beta}^2}}{\Delta f}$	$k_{\mathrm{MIX}} = \dfrac{\overline{v_{n\,\mathrm{MIX\,a}}^2}/\Delta f}{\overline{v_{n\,\mathrm{MIX}\,\beta}^2}/\Delta f}$
LPF	Base amplifier	$\overline{v_{n\,\mathrm{LPF\,a}}^2} = 4k_B T R_{n\,\mathrm{LPF\,a}} \Delta f$	$R_{n\,\mathrm{LPF\,a}}$ is the LPF opamps equivalent noise resistance
	Feedback network	$\overline{v_{n\,\mathrm{LPF}\beta}^2} = 4k_B T R_{n\,\mathrm{LPF}\beta} \Delta f$	$R_{n\,\mathrm{LPF}\,\beta}$ is the LPF feedback network equivalent noise resistance
	TOTAL	$\dfrac{\overline{v_{n\,\mathrm{LPF}}^2}}{\Delta f} = \dfrac{\overline{v_{n\,\mathrm{LPF\,a}}^2}}{\Delta f} + \dfrac{\overline{v_{n\,\mathrm{MIX}\,\beta}^2}}{\Delta f}$	$k_{\mathrm{LPF}} = \dfrac{\overline{v_{n\,\mathrm{LPF\,a}}^2}/\Delta f}{\overline{v_{n\,\mathrm{LPF}\beta}^2}/\Delta f}$
VGA	Base amplifier	$\overline{v_{n\,\mathrm{VGA\,a}}^2} = 4k_B T R_{n\,\mathrm{VGA\,a}} \Delta f$	$R_{n\,\mathrm{VGA\,a}}$ is the VGA opamps equivalent noise resistance
	Feed-back resistors	$\overline{v_{n\,\mathrm{VGA}\,\beta}^2} = 4k_B T R_{n\,\mathrm{VGA}\,\beta} \Delta f$	$R_{n\,\mathrm{VGA}\,\beta}$ is the VGA feedback network equivalent noise resistance
	TOTAL	$\dfrac{\overline{v_{n\,\mathrm{VGA}}^2}}{\Delta f} = \dfrac{\overline{v_{n\,\mathrm{VGA\,a}}^2}}{\Delta f} + \dfrac{\overline{v_{n\,\mathrm{VGA}\,\beta}^2}}{\Delta f}$	$k_{\mathrm{VGA}} = \dfrac{\overline{v_{n\,\mathrm{VGA\,a}}^2}/\Delta f}{\overline{v_{n\,\mathrm{VGA}\,\beta}^2}/\Delta f}$

Hence, the F_{LF} becomes:

$$F_{\mathrm{LF}} = 1 + (1 + k_{\mathrm{LF}}) 8 R_{n\ \mathrm{LPF}\beta} / R_S \tag{14}$$

$R_{n\ \mathrm{LPF}}$, and subsequently $R_{n\ \mathrm{MIX}}$ and $R_{n\ \mathrm{VGA}}$, can be calculates as:

$$R_{n\ \mathrm{LPF}\beta} = R_S \frac{F_{\mathrm{LF}} - 1}{8(1 + k_{\mathrm{LF}})} = R_{n\ \mathrm{MIX}\,\beta}/2 = R_{n\ \mathrm{VGA}\,\beta}/2 \tag{15}$$

Fig. 3.a plots $R_{n\ \mathrm{LPF\ a}}$ and $R_{n\ \mathrm{LPF}\ \beta}$, while Fig. 3.b shows the $R_{n\ \mathrm{MIX\ a}}$, $R_{n\ \mathrm{MIX}\ \beta}$, $R_{n\ \mathrm{VGA\ a}}$ and $R_{n\ \mathrm{VGA}\ \beta}$ as a function of k_{LPF}.

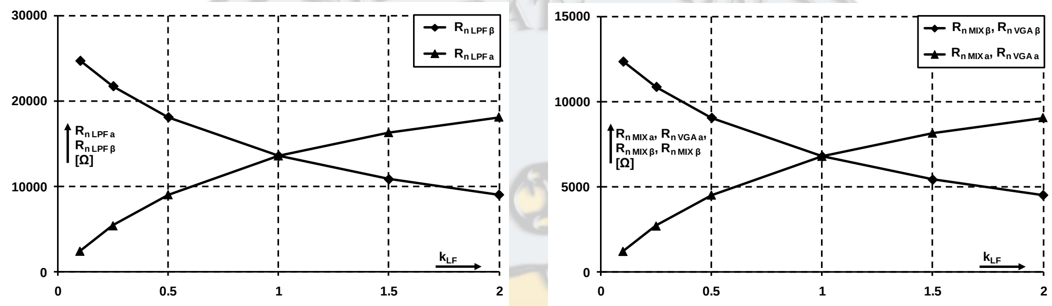

Fig. 3. – a. $R_{n\ \mathrm{LPF\ a}}$, $R_{n\ \mathrm{LPF}\ \beta}$ and b. $R_{n\ \mathrm{MIX\ a}}$, $R_{n\ \mathrm{MIX}\ \beta}$, $R_{n\ \mathrm{VGA\ a}}$, $R_{n\ \mathrm{VGA}\ \beta}$ vs. k_{LPF}.

Finally, the proposed receiver baseband noise partitioning gives the same importance to both area and power consumption by setting $k_{\mathrm{LF}} = 1$. It results, $R_{n\ \mathrm{MIX}\ \beta} = R_{n\ \mathrm{VGA}\ \beta} \leq 7.5\ \mathrm{k}\Omega$ and $R_{n\ \mathrm{LPF}\ \beta} \leq 15\ \mathrm{k}\Omega$.

5. Conclusions

The paper presented the noise partitioning strategy for the low frequency part of a multi standard direct conversion wireless receiver embedding analog signal conditioning.

The noise of the LF chain is split between the noise contributions from the LF operational amplifiers, $\overline{v_{n\ \mathrm{LF\ a}}^2}$, and from the LF part resistances embedded in the feedback network, $\overline{v_{n\ \mathrm{LF}\ \beta}^2}$.

By introducing a factor $k_{\mathrm{LF}} = \overline{v_{n\ \mathrm{LF\ a}}^2} / \overline{v_{n\ \mathrm{LF}\ \beta}^2}$ that measures the LF part excess noise factor due to the baseband operational amplifiers, the LF blocks noise can be referred only to the noise of the opamp feedback network.

Hence, the noise partitioning strategy of the receiver baseband blocks is bounded to the power consumption / area trade-off.

The larger the noise of the base opamps (i. e. larger k_{LF}, equivalent to lower opamp current consumption), a smaller value for feed-back resistors is required to keep the same noise contribution for the baseband blocks; but, in the same time, the integrated capacitance must be increased accordingly to maintain the same LF chain bandwidth.

Finally, the receiver baseband noise partitioning gives the same importance to both area and power consumption by setting $k_{LF} = 1$.

Acknowledgment

The authors would like to express their acknowledgment to Dr. Frank Op't Eynde for triggering the presented analysis and for the fruitful discussions on the topic.

REFERENCES

[1] T. H. Lee, *The Design of CMOS Radio-Frequency Integrated Circuits*, 2^{nd} Edition, Cambridge University Press, 2004, pp. 710-713.

[2] F. Op't Eynde, *A maximally-digital radio receiver front-end*, International Solid-State Circuit Conference, ISSCC 2010, pp. 450-451.

[3] E. Lopelli, S. Spiridon, J.v.d Tang, *A* 40 nm *wideband direct-conversion transmitter with sub-sampling-based output power, LO feedthrough and I/Q imbalance calibration*, International Solid-State Circuit Conference, ISSCC 2011.

[4] S. Spiridon, F. Spiridon, C. Dan, M. Bodea, *An analysis of CMOS re-configurable multi-standard radio receivers building blocks core*, Revue Roumaine des Sciences Techniques, Série Électrotechnique et Énergétique, Issue 1, 2011.

[5] P. Gray, P. Hurst, S. Lewis, R. Meyer, *Analysis and Design of Analog Integrated Circuits*, 4^{th} Edition, Wiley, 2001, pp. 553-612.

[6] S. Spiridon, F. Op't Eynde, *An optimized opamp topology for the low frequency part of a direct-conversion multi-standard radio transceiver*, Proceedings of the First International Symposium on Electrical and Electronics Engineering, Galati, Romania, October 2006, pp. 11-16.

[7] S. Spiridon, F. Spiridon, C. Dan, M. Bodea, *Deriving the key electrical specifications for a multi-standard radio receiver*, The First Intl. Conf. on Advances in Cognitive Radio, COCORA 2011, Budapest.

[8] J. D. Kraus, *Radio Astronomy*, McGraw-Hill, 1966.

[9] S. Spiridon, F. Op't Eynde, *Low power CMOS fully differential programmable low pass filter*, Proc. 10^{th} Intl. Conf. on Optimization of Electrical and Electronic Equipment OPTIM 2006, May 2006, pp. 21-25.

[10] S. Spiridon, F. Op't Eynde, *Low power CMOS fully differential variable-gain amplifier*, Proc. of the Annual Intl. Semiconductor Conf. CAS 2005, October 2005, vol. 2, pp. 383-386.